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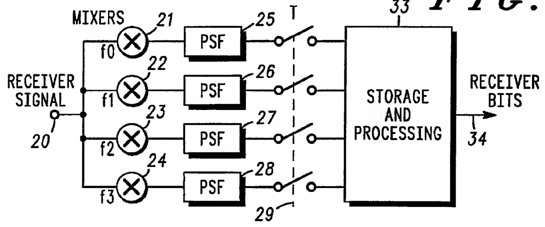
(54) Communications system having pilot signals transmitted over frequency divided channels

(57) In a communications system signals, which include predetermined pilot signals, are transmitted over selected channels of a number of frequency-divided channels, such as in an M-QAM system. The receiver comprises means for receiving predetermined pilot signals simultaneously on selected (e.g. outer) channels of the plurality of channels and for calculating (at 33) channel gain values for said channels, first computation means for calculating frequency domain characteristics (r.m.s. delays) of the channels representative of the relative behaviour of the channels, based on the reception of pilot signals received simultaneously on said selected channels, second computation means for interpolating or extrapolating in the frequency domain the behaviour of channels on which no pilot signal is received in a given timeslot based on the behaviour of channels on which pilot signals are received in that timeslot to provide channel gain values for the remaining channels and third computation means responsive to said channel gain values, for demodulating the signals received on the channels on which no pilot signals are received.

FIG.7



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At least one drawing originally filed was informal and the print reproduced here is taken from a later filed formal copy.

FIG.1

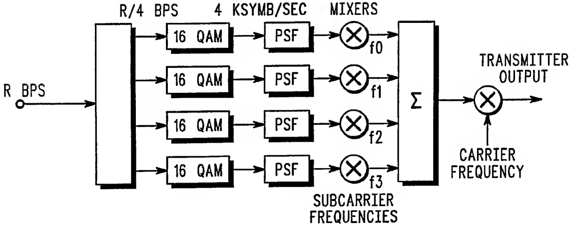


FIG.2

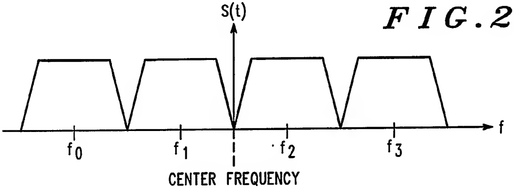


FIG.3

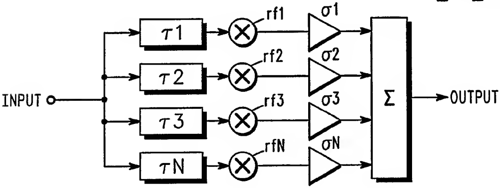
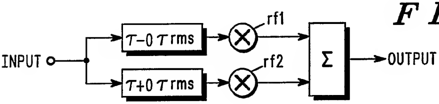


FIG.4



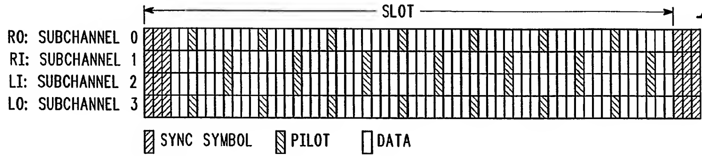
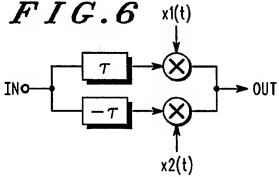


FIG.5

FIG.6



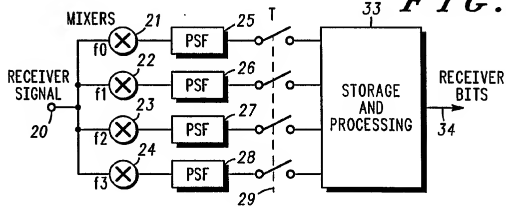


FIG.7

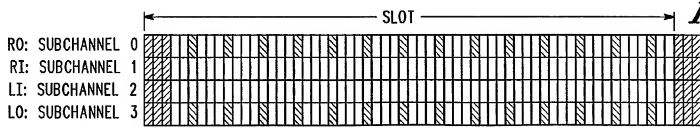
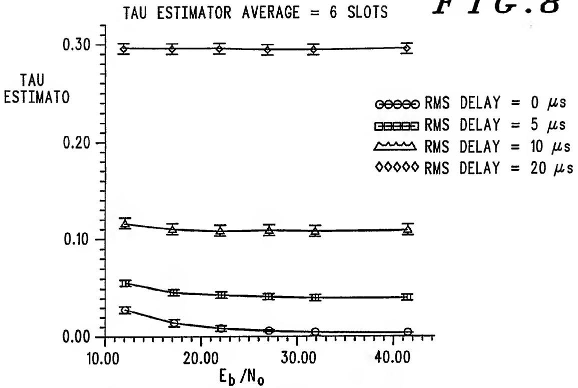


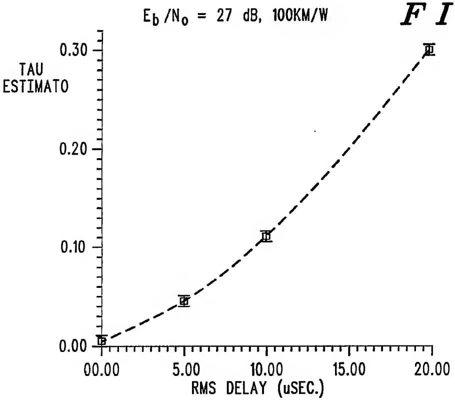
FIG.10

FIG.8



SQUARE OF RMS DELAY ESTIMATOR VS. Eb/No

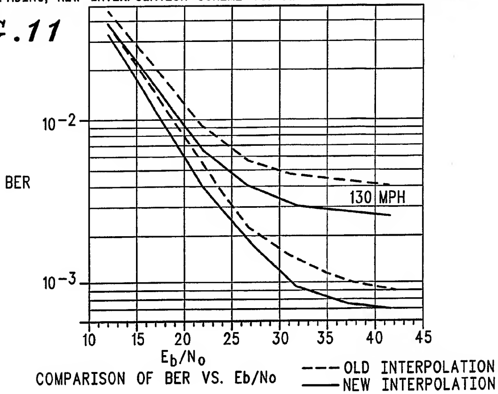
FIG.9



ρ VS. RMS DELAY

FLAT FADING, NEW INTERPOLATION SCHEME VS. OLD INTERPOLATION SCHEME.

FIG.11



COMPARISON OF BER VS. Eb/No

FIG.12

TYPICAL URBAN, 75 MPH μSEC. RMS DELAY

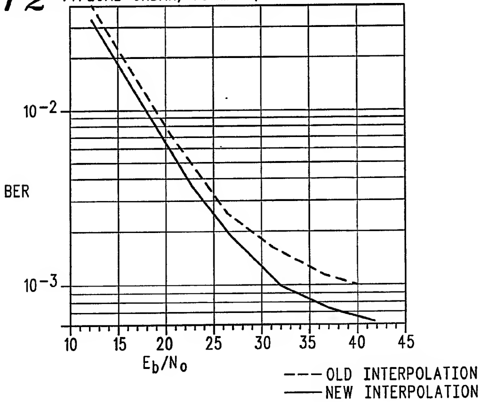


FIG.13

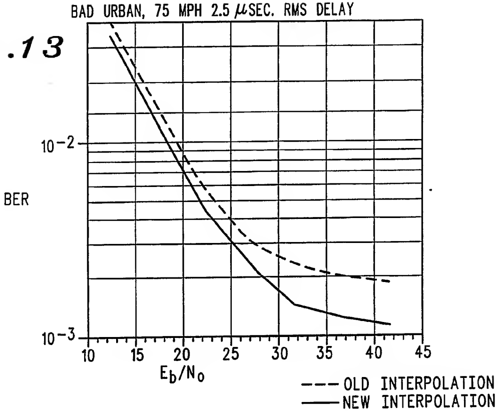
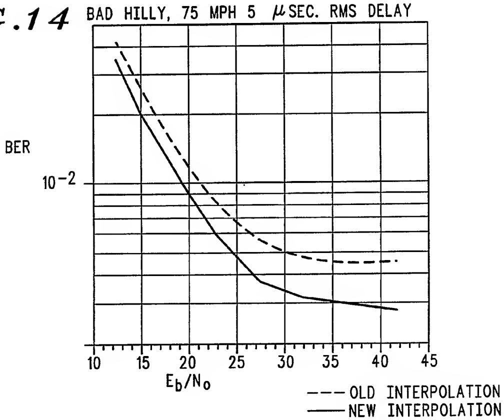

BAD URBAN, 75 MPH 2.5 μSEC. RMS DELAY
BER
10^-2
10^-3
10
15
20
25
30
35
40
45
Eb/No
OLD INTERPOLATION
NEW INTERPOLATION

FIG.14


BAD HILLY, 75 MPH 5 μSEC. RMS DELAY
BER
10^-2
10
15
20
25
30
35
40
45
Eb/No
OLD INTERPOLATION
NEW INTERPOLATION

2271693

Communications System Having Pilot Signals Transmitted Over Frequency Divided Channels

Field of the Invention

This invention relates to a communications systems where signals, which include predetermined pilot signals, are transmitted over selected channels of a number of frequency-divided channels, such as in an M-QAM system.

Background of the Invention

One of the problems associated with highly spectrum efficient digital transmission over land mobile channels is the frequency selectivity of the channel together with its Rayleigh fading characteristics. To overcome this problem, two solutions have been used: adaptive equalization or frequency division multiplexing (FDM). This invention is primarily but not exclusively related to the latter. With this technique, the entire channel (25 KHz for example) is subdivided into subchannels. The subchannels are narrow enough so that the transmission media over this reduced bandwidth can be considered as non-frequency selective or having frequency flat fading (F^3) characteristics. This technique is denoted Multi-Quadrature Amplitude Modulation or M-QAM.

In every subchannel, a sequence of predetermined pilots is embedded in the information in order to compensate for the flat fading introduced by the channel.

A block diagram of a baseband implementation of such a system is depicted in Fig. 1 for the particular case of 4-QAM. The input bit stream at a rate of R bits/sec is subdivided into 4 bit streams at rates R/4 bits/sec. A QAM modulator transforms every 4 bits in the subchannel into a single QAM symbol drawn from a 16QAM constellation (see for example, Proakis, Digital Communications, Mc-Graw Hill). Known symbols termed pilots are inserted in every one of the subchannels at predetermined times NT where T is the symbol rate. The resulting symbol rate for each subchannel in this particular scheme is equal to 4 Ksymbols/sec. The resulting QAM signal is then filtered with a pulse shaping filter (PSF). Each subchannel signal AM modulates an

oscillator centered at the subchannel frequency offset (offset w.r.t. the center frequency of the channel). The resulting signals are then added and translated to the actual transmission frequency. The purpose of the PSF in the subchannel branches is twofold. It limits the spectrum of the subchannel signals in order to avoid interchannel interference and the emission (splatter) into adjacent channels. The spectrum at the output of the transmitter is schematically depicted in Fig. 2.

The signal is transmitted through the land mobile channel. This channel can be characterized as a multiple ray channel where every ray is independently Rayleigh faded having a gain equal to σ_i. A channel model is depicted in Fig. 3. The channel is characterized by its rms delay spread defined as

$$\tau_{rms} = \left[\frac{\sum_{i=1}^{n} (\sigma_i \tau_i)^2}{\sum_{i=1}^{n} \sigma_i^2} \right]^{1/2}$$

(This notation assumes that the average delay is equal to zero or has been subtracted from the individual delays).

Normally, when the rms delay of the channel is not too large with respect to the inverse of the bandwidth of the channel, the multiple ray model can be simplified to a two ray model having the same rms delay spread. This simplified channel is depicted in Fig. 4 where τ_0 is the average channel delay. Each channel is individually Rayleigh faded. The spectrum of the Rayleigh fading signal is characterized by its maximum Doppler frequency f_d given as

$$f_d = \frac{v}{c} f_c$$

where v is the vehicle speed, c is the speed of light and f_c is the actual transmission frequency. For 1.5 GHz and at a speed of some 100 km/h, f_d is equal to 140 Hz. This Rayleigh fading can be modeled in baseband as a complex Gaussian process with in-phase and quadrature components having power spectral densities equal to

$$S_{i/q}(f) = \left[1 - (f / f_d)^2\right]^{-1/2}$$

for $|f| < f_d$ and zero otherwise. This process is a relatively slow varying process with respect to the symbol rate (140 Hz vs. 4 KHz).

The signal at the receiver is contaminated by white Gaussian noise. With an oscillator centered at each subchannel frequency, the signal is split

into its four subchannel components. A low pass filter in each subchannel rejects the signal corresponding to unwanted subchannels as well as noise. This signal is suitably sampled at the symbol rate. The amplitude and phase of the signal in each subchannel needs to be corrected for the variations introduced by the slow varying Rayleigh fading process. This is done in the following way: the gain at the pilot instants is determined by dividing the received signal at the pilot instant by the known pilot symbol. Once gains are known at the pilot instants and since the Rayleigh fading is slow compared to the symbol rate, interpolation can be done in this time domain to derive the gains at every symbol instant. The samples at the symbol instant are then divided by the estimated gain of the channel through this interpolation process and compared to a set of predetermined thresholds to determine the actual QAM symbol transmitted. The symbol is then converted to bits. The bits of the four subchannels are then combined to produce the original transmitted bit stream.

Since the Rayleigh fading process has a certain limited bandwidth, according to the Nyquist criteria, in order to be able to recover this process, samples of this process need to be measured at at least twice the maximum frequency of the process. In the example above, f_c was equal to 140 Hz and pilots ideally need to be included in each stream at a rate of at least 280 Hz, or 280 pilots/sec. This assumes perfect frequency synchronization between transmitter and receiver. Although an automatic frequency control (AFC) circuit is normally implemented in such a system, some residual frequency offset between transmitter and receiver is normally present after AFC correction. In this case, the low pass Rayleigh fading process as seen by the receiver will not be centered around zero but around this frequency offset, increasing the perceived bandwidth at the receiver. To compensate for that, more pilots need to be introduced. Let us assume a residual offset of some 200 Hz. The maximum frequency of the combined process (frequency offset plus Rayleigh fading) will be equal to 340 Hz. Again, using the Nyquist criteria, pilots need to be introduced at a rate larger that 680 Hz. A more real situation where interpolation cannot be accomplished with an ideal brick wall filter, some 800 pilots/sec will be more desirable. This means that one out of every 5 symbols (800 pilots/sec out of 4 ksymbol/sec) in each subchannel is 'wasted' in this correction process, that is 20% of the information transmitted through the channel.

A possible prior art way to reduce this overhead is the following. Let the four subchannels be denoted as left/right (L/R) inner/outer (I/O), i.e., LO, LI, RI, RO. Pilots are introduced alternatively in the inner and outer channel in a way that will be explained below. The information in the channel is send in time slots using 4-16QAM modulation as described above. At the beginning of each slot, a sync pattern is used to aid in the time recovery process and AFC. The transmission in the outbound link is continuous, meaning that each slot is immediately followed by the next slot . The sync pattern consists of 3 predetermined symbols for each subchannel for a total of 16 symbols. Now, instead of sending pilots in each subchannel as described above, they are distributed between inner and outer subchannels. The distribution of pilots in a slot is presented in Fig. 5. The subchannel is sufficiently narrow so that one can assume flat fading (non-frequency selective) for each subchannel. In the time domain interpolation process, the gain at the pilot instants is determined for the inner subchannels and the outer subchannels. To find the gain for any other symbol time say in one outer channel, LO for example, the gain at the pilot instant in the inner channel (LI in this example) together with the gain in the outer channel are taken into consideration. This off-channel gain however cannot be used as if it were an on-channel gain since over the bandwidth of two subchannels, the fading cannot be assumed to be non-frequency selective. To alleviate the problem, in the time domain interpolation process a different weight is given to the off-channel gain at the pilot instants. This reduced weight was optimized through simulations and is constant independent of the rms delay present in the channel. This approach is not in anyway optimum.

Summary of the Invention

According to the present invention, a communications system is provided comprising a transmitter and a receiver, the transmitter comprising means for transmitting signals over a plurality of time slots on a plurality of frequency-divided channels (which in the preferred embodiment are sub-channels of a broader channel), said signals including predetermined pilot signals, wherein the pilot signals are transmitted simultaneously on selected channels of the plurality of channels and the receiver comprising: means for receiving predetermined pilot signals simultaneously on selected channels of

the plurality of channels and for calculating channel gain values for said channels, first computation means for calculating frequency domain characteristics of the channels representative of the relative behaviour of the channels, based on the reception of pilot signals received simultaneously on said selected channels, second computation means for interpolating or extrapolating in the frequency domain the behaviour of channels on which no pilot signal is received in a given timeslot based on the behaviour of channels on which pilot signals are received in that timeslot to provide channel gain values for the remaining channels and third computation means responsive to said channel gain values, for demodulating the signals received on the channels on which no pilot signals are received.

In this manner, frequency domain interpolation or extrapolation is used to improve the receiver operation. The frequency domain characteristics are preferably the r.m.s. delays of the channels (i.e. the sub-channels).

The invention can be employed in a system having pilot signals on all the channels, but it is particularly preferred that the transmitter transmits the pilot signals on those channels that are at the maximum separation in frequency. I.e., in the case of a number of sub-channels of a broader channel, the pilot signals are transmitted only (or substantially only) in the outermost sub-channels. This allows for interpolation between the frequencies (of the sub-channels) rather than extrapolation and provides better results.

A preferred embodiment of the invention is now described, by way of example only, with reference to the drawings.

Brief Description of the Drawings

Fig. 1 shows a block diagram of a typical prior art 16 QAM transmitter;

Fig. 2 shows a schematic representation of the spectrum at the output of the transmitter of Fig. 1;

Fig. 3 shows a channel model for explanation of a typical radio channel;

Fig.4 shows a simplified model of a radio channel;

Fig. 5 shows a prior art time division multiplexed communication channel slot configuration;

Fig. 6 shows a typical two-ray Rayleigh fading model for explanation of the invention;

Fig. 7 shows a block diagram of a receiver for explanation of the present invention;

Fig. 8 shows the results of simulations carried out at different noise levels;

Fig. 9 shows the proposed rms delay estimator as a function of the rms delay in the channel (or subchannel);

Fig. 10 shows a new slot configuration in accordance with a preferred embodiment of the present invention; and

Figs. 11, 12, 13 and 14 show different anticipated bit error rates for different signal-to-noise ratios in different environments.

Detailed Description of a Preferred Embodiment

In at least its preferred embodiment, the present invention requires that the RMS delay in the channel is known or is calculated. A preferred method of estimating the RMS delay is provided by the following mathematical derivation.

Looking first at the typical two ray Rayleigh fading model shown in Fig. 6 where the average delay has been ignored for simplicity. We assume that the first ray is delayed by an amount while the second one is delayed by - where is the rms delay in the channel. Both rays are then independently Rayleigh faded. $x_1(t)$ and $x_2(t)$ represent the two complex Gaussian processes associated with the Rayleigh fading.

The impulse response of the system can be expressed as

$$g(t) = x_1\delta(t-\tau) + x_2\delta(t+\tau)$$

Note that x_1 and x_2 are time dependent.

Taking now the Fourier transform, a time varying frequency response is obtained thus:

$$G_t(f) = x_1 \exp(-\omega\tau) + x_2 \exp(\omega\tau)$$

A sync pattern is sent at the beginning of each slot. During the sync symbols, we can estimate the gain of the channel at four different frequencies, namely the subcarrier frequencies. Note that the time for all the symbols is identical. If we assume negligible noise in this estimation process, we may write

$$G(f_0) = x_1 \exp(-\omega_0 \tau) + x_2 \exp(\omega_0 \tau)$$
$$G(f_1) = x_1 \exp(-\omega_1 \tau) + x_2 \exp(\omega_1 \tau)$$
$$G(f_2) = x_1 \exp(-\omega_2 \tau) + x_2 \exp(\omega_2 \tau)$$
$$G(f_3) = x_1 \exp(-\omega_3 \tau) + x_2 \exp(\omega_3 \tau)$$

where f_0, f_1, f_2, f_3 are the four subcarrier frequencies and $\omega_i = 2\pi f_i$.In this particular case, these subcarrier frequencies are symmetrically spaced around zero and are related as follows.

$$f_0 = -3f_2$$
$$f_1 = -f_2$$
$$f_3 = 3f_2$$

Substitution of this relation in the above equation yields:

$$G(f_0) = x_1 \exp(3\omega_2 \tau) + x_2 \exp(-3\omega_2 \tau)$$
$$G(f_1) = x_1 \exp(\omega_2 \tau) + x_2 \exp(-\omega_2 \tau)$$
$$G(f_2) = x_1 \exp(-\omega_2 \tau) + x_2 \exp(\omega_2 \tau)$$
$$G(f_3) = x_1 \exp(-3\omega_2 \tau) + x_2 \exp(-3\omega_2 \tau)$$

From this last set of equations, there exist several ways of determining the rms delay in the channel. All of them in some way rely on the fact that if the channel is frequency selective, the complex vectors representing the gain will not be collinear. A simple form is derived next.

Let us define

$$\alpha_{02} = G(f_0)G^*(f_2)$$
$$\alpha_{13} = G(f_1)G^*(f_3)$$
$$\alpha_{00} = G(f_0)G^*(f_0)$$
$$\alpha_{11} = G(f_1)G^*(f_1)$$
$$\alpha_{22} = G(f_2)G^*(f_2)$$
$$\alpha_{33} = G(f_3)G^*(f_3)$$

Next define

$$\lambda = \alpha_{11} + \alpha_{22} - \alpha_{02} - \alpha_{13}$$

By direct substitution, it is easily shown that

$$\lambda = 2\left(|x_1|^2 + |x_2|^2\right)(1 - \cos 4\omega_2 \tau)$$

If we denote now

$$\sigma_r^2 = \mathrm{E}|x_1|^2 = \mathrm{E}|x_2|^2$$

where $\mathrm{E}|x_i|^2$is the signal energy in ray i, which in our model is identical for the two channels, then

$$\mathrm{E}(\lambda) = 4\sigma_r^2(1 - \cos 4\omega_2 \tau)$$

Now, if $\omega_2 \tau << 1$we can approximate $\cos(x) = 1 - x^2/2$ and

$$E(\lambda)=32\sigma_r^2\omega_2^2\tau^2$$

that is $E(\lambda)$ is proportional to the square of the rms delay spread. Since σ_r is not really known, let us define

$$\mu=\alpha_{00}+\alpha_{11}+\alpha_{22}+\alpha_{33}$$

It can be easily shown that since $E(x_1x_2^*)=0$ (independent fade assumption) then

$$E(\mu)=8\sigma_r^2$$

We define the ratio

$$\rho=\frac{E(\lambda)}{E(\mu)}=4\omega_2^2\tau^2$$

This ratio is independent of the signal energy in the rays and proportional to the square of the rms delay of the channel.

The above derivation can be generalized to multiple rays as follows. Assume that we have n independently faded rays . The transfer function of the channel can be written as

$$G(f)=\sum_{i=1}^{n}x_i\exp(j\omega\tau_i)$$

We may assume that the clock recovery circuit in the system takes care of the average delay so that

$$\sum_{i=1}^{n}\tau_i=0$$

We define, as above, the set of variables α. We have that

$$\alpha_{02}=\sum_{i=1}^{n}|x_i|^2\exp(-j4\omega_2\tau_i)+\text{cross terms}$$

$$\alpha_{13}=\sum_{i=1}^{n}|x_i|^2\exp(j4\omega_2\tau_i)+\text{cross terms}$$

$$\alpha_{ii}=\sum_{i=1}^{n}|x_i|^2+\text{cross terms}$$

were the term 'cross terms' designates products of the form $x_ix_j^*$ with $i\neq j$.

We define, as above,

$$\lambda=\alpha_{11}+\alpha_{22}-\alpha_{02}-\alpha_{13}$$

and

$$\mu=\alpha_{00}+\alpha_{11}+\alpha_{22}+\alpha_{33}$$

Since we assume independent fades for the rays, the $E(x_ix_j^*)=0$for $i\neq j$. With this condition, it is easily shown that

$$E(\lambda)=2\sum_{i=1}^{n}\sigma_i^2(1-\cos4\omega_2\tau_i)$$

and

$$E(\mu) = 4\sum_{i=1}^{n} \sigma_i^2$$

where

$$\sigma_i^2 = E|x_i|^2$$

represents the energy in every one of the rays.

We define, as above, the ratio

$$\rho = \frac{E(\lambda)}{E(\mu)}$$

$$= \frac{1}{2} \frac{\sum_{i=1}^{n} \sigma_i^2 (1 - \cos 4\omega_2 \tau_i)}{\sum_{i=1}^{n} \sigma_i^2}$$

Assuming now that $\omega_2 \tau_i << 1$ for all i, we can rewrite

$$\rho \approx 4\omega_2^2 \frac{\sum_{i=1}^{n} \sigma_i^2 \tau_i^2}{\sum_{i=1}^{n} \sigma_i^2}$$

which is proportional to the square of the rms delay spread, i.e.,

$$\rho \approx 4\omega_2^2 \tau_{rms}^2$$

It has been shown that this quite simple procedure can help determine the rms delay in the channel. As mentioned, the procedure is based on the fact that the gains for each subchannel are not collinear. Other procedures can be devised that use this same fact.

A simulation program was written to find out for a two ray model the validity of our derivation. A block diagram of the receiver is depicted in Fig. 7. The receiver comprises a receive signal input 20 for receiving a digitized received signal from an R.F. stage (not shown). The received signal is simultaneously processed in four paths by being fed (in software) to four mixers 21 - 25 for injecting four subcarriers and in effect reducing the four components to zero i.f. One pulse shaping filter (PSF) 26 - 29 is provided in each path and the outputs of the PSFs are sampled by sampler 29 every T seconds (i.e. 1/T is the symbol rate). The samples are fed to storage and processing means 33 which stores the samples, calculates the channel gain values as described above, multiplies the samples by the gain values and performs threshold comparisons to provide the received bits, which are output as a received bit stream 34.

Since, in the outbound channel a slot is followed by another slot, the mobile receiver can use in its delay estimation process the gains corresponding to the sync pattern in its own slot and the sync pattern

corresponding to the next slot, for a total of six sync symbols. and are computed for every sync symbol and their quotient is averaged over the six sync symbols. There is a possibility of averaging over more that one received slot to obtain more accurate results.

In the above derivation, it is assumed that the noise associated with the gain estimation process is negligible. To see the effect of the noise, simulations for different cases have been performed. The estimation is averaged over six received slots. Fig. 8 shows the results obtained for as a function of E_b/N_o, the average energy per transmitted bit over the energy per unit of bandwidth of the noise process. The 'one sigma' points are also shown in the graph. It is seen that for E_b/N_o exceeding some 20 dB, the noise has no effect on the estimator.

Fig 9 presents values for as a function of the rms delay. The parabolic nature of the behavior is clearly seen. The small variance of is also evident.

Based on the knowledge of the rms delay spread in the channel, it will be shown that by sending pilots in two of the subchannels, the gain in the other subchannels can be estimated through a straight forward frequency domain interpolation technique. One possible slot structure that uses pilots in two subchannels at a time is depicted in Fig. 10. The same technique can be applied to any other slot configuration that sends two pilots at a time, not only in the case of 4-QAM considered here but for a general M-QAM with M>2.

Observe that, in this example, the pilots have been placed in the outer subchannels. They could have been placed in the inner subchannels or one inner and one outer, etc. This configuration however that uses 'interpolation' in the frequency domain rather that 'extrapolation' offers better results.

Interpolation is done in two steps: frequency domain and time domain. Since the pilots are known to the receiver, the gain at the pilot instants can be determined by dividing the received sample by the corresponding pilot symbol. In this case, we will not assume negligible noise for this gain estimation.

Assume that the gains at the pilot instants for subchannels 0 and 3 can be expressed as

$$\hat{g}_0 = g_0 + \eta_0$$
$$\hat{g}_3 = g_3 + \eta_3$$

where g_i is the actual gain in subchannel i while η_i is the 'measurement noise'. The problem can be postulated as follows: Find a_0 and a_3 such that

$$\hat{g}_1 = a_0\hat{g}_0 + a_3\hat{g}_3$$

and

$$\Gamma = \mathrm{E}\left|\hat{g}_1 - g_1\right|^2 \to \min$$

Direct substitution yields

$$\hat{g}_1 - g_1 = a_0(g_0 + \eta_0) + a_3(g_3 + \eta_3) - g_1$$

From the two ray model, we can write

$$g_0 = x_1 \exp(-j3\omega_2\tau) + x_2 \exp(j3\omega_2\tau)$$
$$g_1 = x_1 \exp(-j\omega_2\tau) + x_2 \exp(j\omega_2\tau)$$
$$g_2 = x_1 \exp(j3\omega_2\tau) + x_2 \exp(-j3\omega_2\tau)$$

With the aid of this last equation, we can compute

$$\begin{aligned}\Gamma = \mathrm{E}\left|\hat{g}_1 - g_1\right|^2 &= a_0^2(2\sigma_r^2 + v^2) + a_3^2(2\sigma_r^2 + v^2)\\ &+4a_0a_3\sigma_r^2\cos(6\omega_2\tau) - 4a_0\sigma_r^2\cos(2\omega_2\tau)\\ &-4a_3\sigma_r^2\cos(4\omega_2\tau) + 2\sigma_r^2\end{aligned}$$

where

$$v^2 = \mathrm{E}\left|\eta_i\right|^2 \quad \text{for } i = 1,2$$

is the power of the noise associated with the gain measurement process.

In order to find the minimum with respect to a_0 and a_3, we differentiate and equate to zero

$$\frac{\partial\Gamma}{\partial a_0} = 0 \Rightarrow a_0(2\sigma_r^2 + v^2) + 2a_3\sigma_r^2\cos(6\omega_2\tau) = \sigma_r^2\cos(2\omega_2\tau)$$
$$\frac{\partial\Gamma}{\partial a_3} = 0 \Rightarrow 2a_0\sigma_r^2\cos(6\omega_2\tau) + a_3(2\sigma_r^2 + v^2) = \sigma_r^2\cos(4\omega_2\tau)$$

A system of two equations in two unknowns has to be solved. The solution to this system of equation is

$$a_0 = \frac{\cos 2\omega_2\tau - \cos 6\omega_2\tau \cos 4\omega_2\tau + \chi^{-1}\cos 2\omega_2\tau}{\sin^2 6\omega_2\tau + 2\chi^{-1} + \chi^{-2}}$$
$$a_1 = \frac{\cos 4\omega_2\tau - \cos 6\omega_2\tau \cos 2\omega_2\tau + \chi^{-1}\cos 4\omega_2\tau}{\sin^2 6\omega_2\tau + 2\chi^{-1} + \chi^{-2}}$$

where we have defined

$$\chi = \frac{2\sigma_r^2}{v^2}$$

which is related to the signal-to-noise ratio in estimating g_0 or g_3. Optimal frequency domain interpolation coefficients have been derived for this case. From symmetry considerations, we can write the following equations for the two cases

$$\hat{g}_1 = a_0\hat{g}_0 + a_3\hat{g}_3$$
$$\hat{g}_2 = a_3\hat{g}_0 + a_0\hat{g}_3$$

It has been shown that from the sync part of the slot, the rms delay in the channel can be estimated. In the two outer channels, gains are estimated using the known pilots and sync symbols. Known sync symbols are also used in the inner channels to estimate the gain at the two ends of the slot (in the outbound case), while in the middle of the slot, frequency domain interpolation can be used.

Once the gains are known at all the corresponding pilot instants, time domain interpolation can be done. To derive time domain interpolation coefficients, several techniques can be used. The simplest is to use a low pass interpolating filter with predetermined coefficients.

Results for several typical environments characterized by different rms delay spreads obtained with this new technique are presented in Fig. 10 to 14 and compared with the suboptimum technique described in the previous section. The center frequency of the system for all cases is assumed to be equal to 900 MHz in both cases. It can be observed there that at a 1\% Bit Error Rate, typical of a mobile communication system, an improvement of at least 1 dB can be obtained using this new technique compared to the old one.

One possible variant of this new technique would be to assume a certain constant rms delay in the channel. In this case, frequency domain and time domain interpolation equations can be combined into a single set of equation for the inner channels.

Although the above examples have emphasized the 4-16QAM technique, it should be clear the this same technique can be used for other multiple modulation schemes using this type of FDM technique with pilot insertion to correct for the unknown gain introduced by the frequency selective Rayleigh fading channel. As examples, all classes of M-NQAM or M-NPSK van be considered where M stands for the number of FDM subchannels and N for the number of points in the quadrature amplitude modulation (QAM) or phase shift keying (PSK).

Claims

1. A communications system comprising a transmitter and a receiver,

the transmitter comprising means for transmitting signals over a plurality of time slots on a plurality of frequency-divided channels, said signals including predetermined pilot signals, wherein the pilot signals are transmitted simultaneously on selected channels of the plurality of channels and

the receiver comprising:

means for receiving predetermined pilot signals simultaneously on selected channels of the plurality of channels and for calculating channel gain values for said channels,

first computation means for calculating frequency domain characteristics of the channels representative of the relative behaviour of the channels, based on the reception of pilot signals received simultaneously on said selected channels,

second computation means for interpolating or extrapolating in the frequency domain the behaviour of channels on which no pilot signal is received in a given timeslot based on the behaviour of channels on which pilot signals are received in that timeslot to provide channel gain values for the remaining channels and

third computation means responsive to said channel gain values, for demodulating the signals received on the channels on which no pilot signals are received.

2. A communications system according to claim 1, wherein the transmitter transmits the pilot signals on selected channels only and no pilot signals on the other channels.

3. A communications system according to claim 2, wherein, of the plurality of channels, the selected channels are at the maximum separation in frequency.

4. A communications system according to claim 1, 2 or 3, wherein the frequency domain characteristics are the r.m.s. delays of the frequency divided channels.

5. A communications receiver for receiving signals over a plurality of time slots on a plurality of frequency-divided channels, comprising:

means for receiving predetermined pilot signals simultaneously on selected channels of the plurality of channels and for calculating channel gain values for said channels,

first computation means for calculating a frequency domain characteristic of the channels representative of the relative behaviour of the channels, based on the reception of pilot signals received simultaneously on said selected channels,

second computation means for interpolating or extrapolating in the frequency domain the behaviour of channels on which no pilot signal is received in a given timeslot based on the behaviour of channels on which pilot signals are received in that timeslot to provide channel gain values for the remaining channels and

third computation means responsive to said channel gain values, for demodulating the signals received on the channels on which no pilot signals are received.

6. A communications receiver according to claim 5 further comprising fourth computation means for interpolating in the time domain the behaviour of all the channels based on the behaviour of channels on which sequential pilot signals are received at sequential times and based on frequency domain characteristics of the channels at said sequential times of

reception of the sequential pilot signals, thereby calculating channel gain values for all the channels in all time slots.

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Examiner's report to the Comptroller under Section 17 (The Search Report)

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Search Examiner

K WILLIAMS

Databases (see over)

(i) UK Patent Office

(ii) ONLINE DATABASE: WPI

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Documents considered relevant following a search in respect of claims 1 TO 6

Category (see over)	Identity of document and relevant passages	Relevant to claim(s)
A	WO 91/20142 A1 (MOTOROLA) see page 3, lines 15-23	1,5
A	WO 91/20140 A1 (MOTOROLA) see page 6, lines 13-17	1,5
X	WO 91/20137 A1 (MOTOROLA) see page 5, lines 20-28; page 10 lines 1-16	1,2,5,6

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SF2(p)

Category	Identity of document and relevant passages	Relevant to claim(s)

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